

## An Adaptive Channel Estimation Scheme for DS-CDMA Systems

P.D. 24/09/2000	5
p. 2839-2843	

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**Abstract**

*We propose using decimation to estimate vehicle speed for adaptive channel estimation in pilot-aided transmission systems like DS-CDMA ones. It is shown that by passing the received pilot signal through bandpass filter after decimation and computing its output energy, current velocity of mobile station can be accurately estimated. In addition, the proposed speed estimation unit is simple to implement. This allows us to use it accompanying computationally efficient pilot filter for forward link receiver. a structure for adaptive channel estimation based on the proposed velocity estimation unit is addressed. The optimal coefficients of pilot filter is selected from a precalculated bank of coefficients based on the estimated mobile speed zone. Procedures for deriving the optimal coefficients and thresholds for the speed estimation unit are developed. The proposed method provides up to 0.75dB improvement compared to conventional non-adaptive ones.*

**1. Introduction**

In the mobile communication systems employing pilot tone as a phase and amplitude reference for demodulation on fading channels, the pilot filter is used to extract the reference. For example, a bank of lowpass filters is employed in DS-CDMA system to estimate the channel response for each finger prior to rake combining. The receiver extracts the pilot using pilot filters with *fixed* bandwidth. In the forward link, some computationally efficient pilot filters like a moving average filter and a first-order IIR filter are preferred for simple implementation. The bandwidth of pilot filter should be wide enough to accommodate the fade spectrum (the maximum Doppler frequency  $f_D$ ). If it is too narrow, the estimated channel response cannot trace the actual channel fluctuation which leads to the appearance of an error floor. In contrast, if it is too wide excessive noise

\* This work was partially supported by Korea Science and Engineering Foundation (KOSEF) and Samsung Telecommunications.

appears in the output of pilot filter. The optimal pilot filter for a given  $f_D$  can be obtained using Wiener filter theory. However, the Wiener filter for a fixed  $f_D$  cannot be optimal in practice because  $f_D$  varies as the mobile speed changes.

The performance penalty caused by fixed bandwidth can be reduced by introducing adaptivity to pilot filtering so that the bandwidth of pilot filter is adjusted to changes in the Doppler. Some adaptive filtering schemes were previously proposed [1]-[2]. Use of the LMS algorithms based on multi-dimensional adaptation [1] is too complex to be used in the forward link. More reliable adaptation algorithm proposed in [2]. When the magnitude response of a pilot filter is not flat in passband like a moving average filter and a first-order IIR filter, it is not adequate. Many literatures have shown to estimate velocity of mobile and corresponding  $f_D$  [3]-[7]. These methods are computationally inefficient or unreliable to be used in practical systems. For example, the performance of the method based on level crossing detector [7] is severely degraded under AWGN.

Another performance penalty of the currently used channel estimation method in DS-CDMA systems is caused by using the prediction-type channel estimator—group delays caused by pilot filtering is not compensated in traffic channel decoding. Compensation of this kind of group delays degrades the performance of power control due to the additional delays in its loop.

In this paper, a new mobile speed estimation method using decimation is introduced. This estimation unit is reliable and simple to implement. Using the proposed mobile speed estimation method, an adaptive channel estimation scheme reducing or eliminating the performance penalties mentioned above is addressed. Up to 0.75dB gain was gained using the proposed scheme.

**2. System model**

In this paper, we only consider the pilot channel for channel estimation in DS-CDMA system. The performance of channel estimation could be improved using not only pilot channel signal but also traffic channel signal. Then trans-

mitted signal  $s(t)$  assumed to be QPSK modulated and QPSK spread is given by

$$\begin{aligned} s(t) = & \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) - a_Q(t)\} \\ & + j \left[ \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) + a_Q(t)\} \right] \\ & + \text{traffic channel signals} \end{aligned} \quad (1)$$

where,  $E_{c,p}$  is pilot signal energy per pseudo noise (PN) chip sequence, and  $a_I(t)$  and  $a_Q(t)$  is I-channel and Q-channel PN sequence, respectively. In Eq. (1), transmitter filter  $h(t)$  is absorbed in  $a_I(t)$  and  $a_Q(t)$ . To simplify analysis, we shall initially assume frequency non-selective fading and single user environment, then received pilot signal  $r(t)$  can be written as

$$\begin{aligned} r(t) = & \frac{1}{2} R_h(\tau) \alpha \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) - a_Q(t)\} \cos \phi \\ & + \frac{1}{2} R_h(\tau) \alpha \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) + a_Q(t)\} \sin \phi \\ & + j \left[ -\frac{1}{2} R_j(\tau) \alpha \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) - a_Q(t)\} \sin \phi \right. \\ & + \left. \frac{1}{2} R_h(\tau) \alpha \sqrt{\frac{E_{c,p}}{2}} \{a_I(t) + a_Q(t)\} \cos \phi \right] \\ & + \text{terms bearing traffic channel signals} \\ & + \frac{1}{2} n_I(t) + j \frac{1}{2} n_Q(t) \end{aligned} \quad (2)$$

where,  $\alpha$  and  $\phi$  is amplitude and phase response of Rayleigh fading, respectively,  $R_h(\tau)$  is autocorrelation function of  $h(t)$  under the assumption of using matched filter at receiver, and  $\tau$  is timing error.  $n_I(t)$  and  $n_Q(t)$  are white Gaussian noise, and their single sided spectral densities are  $N_o$ 's. Under the assumption that there is no timing error ( $\tau = 0$ ), the terms bearing traffic channel signals in Eq. (2) can be discarded. Moreover,  $R_h(0) = 1$ . Finally, pilot despread received signal  $y_p(n)$  prior to pilot filtering is expressed as

$$\begin{aligned} y_p(n) = & \alpha(n) N \sqrt{2E_{c,p}} \cos \phi(n) \\ & + j \alpha(n) N \sqrt{2E_{c,p}} \sin \phi(n) \\ & + n'_I(n) + j n'_Q(n) \end{aligned} \quad (3)$$

where,  $N$  is a spreading factor and  $E[n'_I(n)^2] = E[n'_Q(n)^2] = NN_o$ . The channel estimation is performed by manipulating  $y_p(n)$  given in Eq. (3) on symbol rate.

### 3. Proposed architecture

The proposed adaptive pilot filtering scheme is based on the estimation of mobile speed. Tap coefficients of pilot filter are updated to adjust its bandwidth to accommodate  $f_D$

corresponding to the estimated speed. Determination of the type and order of pilot filter  $p(z)$  depends on applications. However, it should be simple to implement for forward link receiver. Throughout this paper, we assume that the following first-order IIR filter is used for pilot filtering—it is more relevant to use a first-order IIR filter for adaptation than to use a moving average filter

$$p(z) = \frac{a}{1 - (1-a)z^{-1}} \quad (4)$$

where,  $a$  is a power-of-two coefficient so that it is simple to implement. Then, it is the only  $a$  to be adaptively updated. In addition, the speed of mobile stations doesn't need to be estimated exactly because only a few discrete values of  $a$  is available. The mobile speed to be estimated is divided into a few number of target speed zones, and a set of the optimal coefficient  $a$  for each speed zone is precalculated and stored. In results, each speed zone has a different pilot filter with a different bandwidth (coefficient  $a$ ) optimized for each velocity. The best one is automatically selected according to the estimated speed zone. The proposed architecture is shown in Fig. 1. The group delay  $d_t$  caused by pilot filtering is compensated for traffic channel decoding. In other hands, a separate path without this compensation is provided for power control. In general, it is impossible to determine the exact value of  $d_t$  because IIR filter has non-constant group delay. In this paper, the group delay at DC component is selected for  $d_t$ . The performance degradation caused by this approximation is negligible. It can be easily shown that  $d_t = \lfloor 1/a \rfloor - 1$ .

#### 3.1. Estimation of mobile speed

The proposed mobile speed estimation method shown in Fig. 2 is based on searching  $f_D$  by using decimation processing. As mentioned above, only a few number of speed zones are sufficient to accommodate a finite number of discrete value of  $a$ . Each branch with a  $M$ -to-1 decimator in Fig. 2 is corresponding to one mobile speed zone. The aliasing caused by decimation is reduced by passing output of pilot filter to  $M$ -to-1 decimator. By decimating the pilot signal by  $M$ , its effective bandwidth is widened by a factor of  $M$ . In results, there exists some  $M'$ 's which make the averaged power of bandpass filtered pilot signal  $e_Q(n)$  larger than the predetermined threshold  $t_i$  as shown in Fig. 2, where integer  $i$  represents each branch. The bandpass filter with high Q-factor here is a simple device for energy monitoring. The current speed zone of mobile station is the one corresponding to the minimum value of  $M'$ . Accordingly, the decimation in Fig. 2 should be continued until  $e_Q(n)$  becomes larger than  $t_i$  at  $i^{th}$  branch for the first time. If there is no  $M$  making  $e_Q(n)$  larger than any  $t_i$ 's, the current speed zone is determined to the slowest velocity. In

fact, one more threshold  $t'_i$  exists at each branch to ensure backward estimation. That is, the current speed zone moves up to faster speed zone of  $(i-1)^{th}$  branch if  $e_Q(n) > t'_i$ . The details for deriving the optimal threshold will be addressed later.

To illustrate the proposed mobile speed estimation unit, consider the following DS-CDMA example: chip rate is 3x1.2288MHz, data rate is 9600bps, symbol rate is 14400sps, processing gain is 256, and carrier frequency is assumed at 2GHz. The center frequency of the bandpass filter in Fig. 2 is fixed at 0.1. Let the current speed of mobile be 50 km/h, then the maximum Doppler frequency  $f_D$  is 92.59Hz which is corresponding to  $f_D = 0.0064$  in normalized frequency. After 8-to-1 decimation,  $f_D$  becomes 0.0514. Because  $f_D \ll 0.1$ ,  $e_Q(n) \ll t_0$ . Therefore, decimation should be continued. After the next 10-to-1 decimation ( $M = 10$ ),  $f_D = 0.0643$  which leads to  $e_Q(n) \ll t_1$  again. After decimating by 16 totally,  $f_D = 0.1029$  and  $e_Q(n)$  becomes larger than  $t_2$  for the first time. So, the estimated current velocity is between 30 km/h and 60 km/h. As velocity increases to 70 km/h,  $e_Q(n)$  becomes larger than  $t'_2$ , so that current velocity is set to the desired speed zone ( $i = 1$ ). In the same way, the last zone is selected as the speed decreases to 20 km/h as  $e_Q(n) < t_2$ . Choosing the number of decimation branches and a decimation ratio in each branch depends on applications.

As total power of received signal increases,  $e_Q(n)$  can be also indefinitely increased, so that thresholds seem to be meaningless any more. In practice, however, it does not happen because automatic gain control (AGC) is employed in every cellular system where dynamic range of received signal is about 80dB. Otherwise, it can be also avoided by normalization.

### 3.2. The optimal pilot filter

$y_p(n)$  in Eq. (3) can be rewritten as

$$\begin{aligned} y'_p(n) &= \frac{1}{N\sqrt{2E_{c,p}\sigma_\beta^2}} y_p(n) \\ &= \frac{1}{\sigma_\beta} \beta(n) + \frac{1}{N\sqrt{2E_{c,p}\sigma_\beta^2}} \tilde{n}(n) \end{aligned} \quad (5)$$

where,  $\beta(n) = \alpha(n) \cos \phi(n) + j\alpha(n) \sin \phi(n)$ ,  $\sigma_\beta^2$  is the variance of  $\beta(n)$ , and  $\tilde{n}(n) = n'_I(n) + jn'_Q(n)$ . The spectrum of  $\beta(n)$  is given by

$$S_\beta(f) = \frac{\sigma_\beta^2}{\pi\sqrt{f_D^2 - f^2}} \quad (6)$$

and its autocorrelation function is

$$R_\beta(\tau) = \sigma_\beta^2 J_0(2\pi f_D \tau). \quad (7)$$

$J_0(\cdot)$  is Bessel function of the first kind with zero order. The optimal coefficient  $a$  of pilot filter  $p(n)$  is obtained by minimizing the following one dimensional optimization problem with respect to  $a$ .

$$\begin{aligned} \min_a \quad & F(a) \\ &= E[|p(n) * y'_p(n) - \beta(n - d_t)|^2] \\ &\approx \sum_k \sum_l p(k)p(l) J_0(2\pi f_D(k-l)T) \\ &\quad + J_0(0) + \frac{N_o}{NE_{c,p}\sigma_\beta^2} \sum_k p(k)^2 \\ &\quad - 2 \sum_k p(k) J_0(2\pi f_D(k - d_t)T) \end{aligned} \quad (8)$$

where,  $T$  is the symbol duration and  $d_t$  is the group delay explained in Section 3. This optimization problem can be easily solved numerically because only a few values of  $a$  are available ( $a$  is a power-of-two coefficient). For instance, if  $2^{-1}$ ,  $2^{-2}$ , and  $2^{-3}$  are available for the value of  $a$ , the optimal one for given  $f_D$  can be obtained by evaluating and comparing the value of  $F(a)$  each other. Notice that the optimal coefficient  $a$  is a function of pilot signal-to-interference ratio (PSIR)  $E_{c,p}\sigma_\beta^2/N_o$  as it is expected.

### 3.3. Determination of threshold $t_i$ and $t'_i$

The pilot filtered  $y_p(n)$  is

$$\tilde{y}_p = N\sqrt{2E_{c,p}}\beta(n) * p(n) + \tilde{n}(n) \quad (9)$$

where,  $E[\tilde{n}^2(n)] = 2NN_o \sum_k p(k)^2$ . Consider  $i^{th}$  branch with M-to-1 decimator. This branch is corresponding to the mobile speed of between  $v_i^l$  and  $v_i^h$ . After M-to-1 decimation and bandpass filtering,  $\tilde{y}_p$  becomes

$$MN\sqrt{2E_{c,p}}\beta(Mn) * p(Mn) + \tilde{n}(n) \quad (10)$$

where,  $b(n)$  is the impulse response of bandpass filter in Fig. 2. We assume that the magnitude response of  $p(Mn)$  in Eq. (10) is almost flat in passband  $[0, f_D]$  and  $\tilde{n}$  is white for simplicity. Then, the total output energy is given by

$$\begin{aligned} e_Q(f_D) &\approx 2M^2N^2E_{c,p}\sigma_\beta^2 \cdot \\ &\quad \sum_k \sum_l b(k)b(l) J_0(2\pi M f_D(k-l)) \\ &\quad + 2MNN_o \sum_k p(k)^2 \sum_k b(k)^2 \end{aligned} \quad (11)$$

In results,  $t_i = e_Q(f_D^l)$  and  $t'_i = e_Q(f_D^h)$ , respectively, and  $f_D^l = v_i^l/\lambda$  ( $f_D^h = v_i^h/\lambda$ ). AGC or normalization can be used for preventing  $e_Q(f_D)$  to increase so large as mentioned Section 3.1.

### 3.4. Practical considerations

Despite all  $a$ ,  $t_i$ 's, and  $t_i'$  are functions of PSIR, they are quite robust to it. It seems to need a huge table of them for every possible PSIR, but only a few number of PSIR's are sufficient because a few number of possible  $a$ 's are available. Roughly estimated PSIR is also accessible in the receiver.

### 4. Simulation results

To demonstrate the efficiency of the proposed method, the proposed method was applied to the forward link of the DS-CDMA system considered in Section 3.2. Convolutional code with the rate of  $1/3$  and the constraint length of 9 is employed. Fast power control was applied with the rate of 800 Hz and power control error of 10%. 20% of the cell power is assigned to the pilot. Jakes' model was used for generating frequency selective Rayleigh fading channel. The first order IIR pilot filter is employed for pilot filtering and the coefficient  $a$  is updated frame by frame. The mobile speed is divided into five speed zones of  $\{[0, 30 \text{ km/h}), [30, 60 \text{ km/h}), [60, 90 \text{ km/h}), [90, 120 \text{ km/h}), \text{ and } [\geq 120 \text{ km/h})\}$  as shown in Fig. 2. Corresponding stored coefficients  $a$  are  $\{16, 8, 4, 4, 2\}$  with  $E_{c,p}/N_o = -13\text{dB}$ . Notice that  $a$ 's in 3<sup>rd</sup> and 4<sup>th</sup> zone are same. We could merge two zones but we don't for the proper operation in the different PSIR's.

In general, frame error rate less than 2% is acceptable for voice decoding. Gains in the averaged transmitted power to achieve the frame error rate less than 2% over conventional one with a fixed  $a = 1/4$ , are shown in Fig. 3 for several vehicle speed in  $\text{km/h}$ . Up to 0.75dB gain (yielding up to 18.7% of capacity increase) is obtained by introducing the adaptivity in channel estimation. The gain drastically increases as the number of multi-path or the allowable complexity in pilot filter increases. The worst case convergence time is occurred when the velocity is between 0  $\text{km/h}$  and 30  $\text{km/h}$ , because it needs more than two frames to get a reliable  $e_Q(n)$ —as decimation ratio  $M$  increases, a few number of symbols in a frame are available for computing it. However, the performance degradation is negligible because the amount of variation in vehicle speed for two or three frames is very small even if an acceleration of 0-100  $\text{km/h}$  in 5 seconds. The details of analysis is not addressed here because of space limitation.

### 5. Conclusion

An adaptive channel estimation method relying on estimating the velocity of mobile station for pilot-aided transmission systems was developed. It was shown that the sig-

nificant improvement in system performance could be provided by introducing adaptivity in pilot filtering. Moreover, the proposed architecture was simple for implementation, so that it was appropriate for forward-link receiver.

### Acknowledgments

The authors would like to thank Dr. Hyung-rae Park and Dr. Beop-ju Kang in Electronics and Telecommunication-s Research Institute (ETRI) for valuable suggestions and comments.

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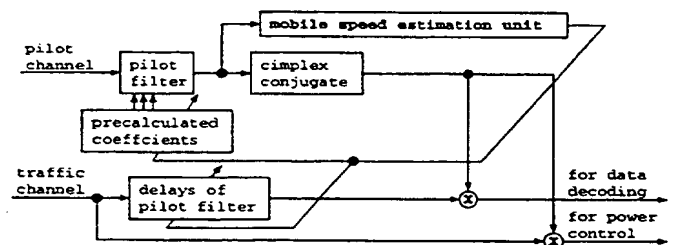


Figure 1. Proposed architecture for channel estimation in DS-CDMA systems.

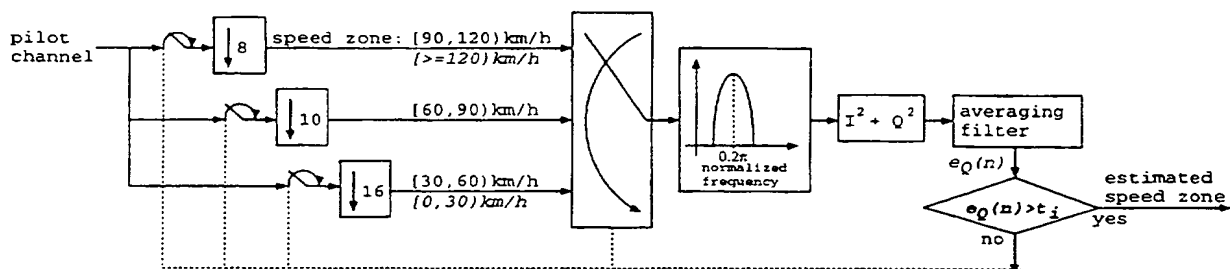


Figure 2. Proposed mobile speed estimation unit.

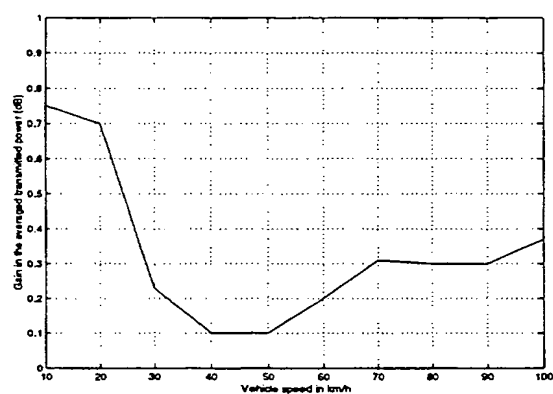


Figure 3. Gain in the averaged transmitted power.

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